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Impact of control Strategy on Component Ratings of Series Resonant DC Link Current Converter

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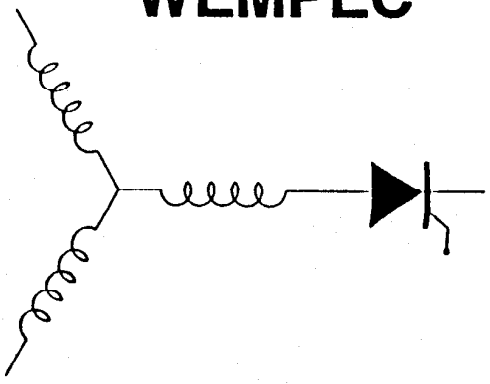

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Abstract

In this paper the voltage and current stresses of a Series Resonant DC Link (SRDCL) Converter are investigated for two improved control strategies, a maximum power transfer method, and a proposed adjacent state vector control method. It is demonstrated that the new method reduces the worst case voltage and current stresses dramatically by only modifying the control strategy and without any sacrifice from the other features of the system such as nearly sinusoidal input currents and unity input power factor. In addition, basic operation principles of the converter are given and some simulation results along with experimental results verifying the theory are presented.

1 Introduction

Switching losses due to hard switching, are often listed as a major obstacle for realizing a controllable dc voltage from an ac source operating at unity power factor. Some control approaches to solve this problem, such as [1] use one or more extra switches which merely trades reduced switching losses for increased conduction losses. Other approaches [2,3] seek to eliminate the switching losses without substantial additional conduction losses by allowing the switching devices to switch only at zero voltage or zero current crossing of the ac input. In particular, a high frequency series resonant dc link (SRDCL) converter, shown in Fig. 1, has recently been developed having the additional advantage of short circuit proof capability. Recently this converter was proposed as a conditioning system for a superconducting energy storage system (SMES) [4] and as a power converter for a dc motor drive [5].

One important problem which remains concerning the operating behavior of the SRDCL converter is the voltage and current stresses on the resonant components due to changing conditions in the load [6]. Dual-flow pulse trimming and current clamp concepts have been recently introduced to help solve this problem [7-9]. However, the extra cost and component losses associated with these modifications are distinct disadvantages.

This paper proposes adjacent state vector control strategy [10] as an alternative to reduce the voltage and current stresses in the resonant capacitor and components. The modification involves only a change in the control strategy. This new strategy has been applied for use in a dc motor drive and the experimental results before and after implementation of the modified control strategy are illustrated and discussed in this paper.

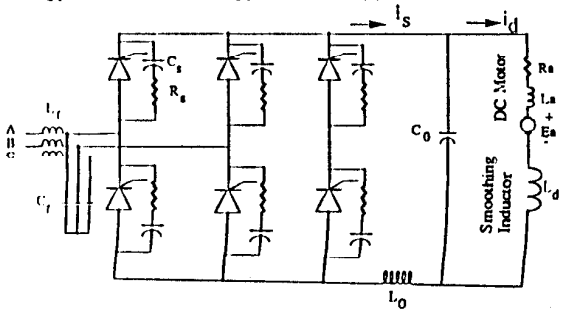

Fig.1 Complete Series Resonant DC Link Converter utilized as a dc motor drive

2 Series Resonant DC Link Current Converter

The SRDCL converter is shown in Fig. 1. By turning on one switch from the upper part and one from the lower part of the converter a voltage is applied to the link which, in turn, creates a resonant circuit consisting of L_0 and C_0. The resonant frequency of the system is about 20 kHz. When the resonant current pulse i_s reaches zero, the switching devices can be readily turned off thereby eliminating the switching losses of the converter. The next pair of the switches are turned on when the voltage across them are equal to a predetermined voltage called V_{swt}. Selection of this quantity is very important and has to be calculated by considering the switch turn-off times, and its effect on the device ratings and operating frequency. Another very important issue concerns the current regulation. The dc current i_d is approximately the average value of the resonant current i_s, and must be regulated so that the resonant current pulses will reach zero during the next resonant interval. To realize a smoother dc current, a smoothing inductance is inserted to the circuit in parallel with the resonant capacitor, and in series with the dc motor. By using a proper control scheme it is possible to obtain sinusoidal input currents and unity power factor at the input terminals.

It is clear that the reduced switching losses, and the capability of sinusoidal input currents with unity input power factor are features of this converter which make this concept very effective and useful. The SRDCL converter can be used to control a superconducting magnetic storage element (SMES) by connecting the SMES in parallel with the resonant capacitor [4]. The SMES can be energized during the low energy demand periods and this energy can be used when there is a demand. During the charging and discharging periods it is possible to maintain unity power factor at the input.

It is also possible to drive an induction machine by the SRDCL converter by connecting a second identical bridge to the link [8]. The resonant pulses are distributed to the stator phases by using Pulse Density Modulation, and by utilizing proper control method, sinusoidal motor currents are obtained along with speed and torque control.

Figure 2 shows the overall control scheme of the SRDCL converter for a dc motor drive. The reference motor current is obtained from the speed error. By using a current regulator and limiter, the current amplitude demand I_{in} is calculated. This variable is multiplied with three-phase sinusoidal reference signals and these yield the references for the input currents. The outputs of the current regulator are used to determine which switches should be turned on next.

One of the current regulation methods at this point, which will be examined in this paper, aims to obtain maximum power transfer by applying to the link the most positive voltage when more current is needed, or most negative voltage when less current is needed. Another means is to use the adjacent state vector control method in which there exists a zero voltage state in each transition from the most positive voltage to most negative voltage, or vice versa [10].

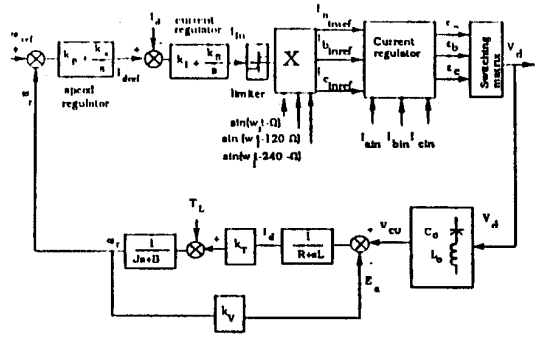

Fig.2 Overall closed loop control scheme

This algorithm will be shown to substantially reduce the voltage and current stresses of the components.

In order to understand how the converter functions, it is practical to use its monophase equivalent circuit as shown in Fig. 3. The fact that the resonant phenomena occurs much faster than the sinusoidal input voltage variations enables one to assume that converter output is a dc voltage source during the period of investigation. Therefore the converter output voltage $v_d(t)$ is assumed to have 3 possible values, namely V_d, 0, $-V_d$. The voltage V_d is the instantaneous line-to-line voltage value, and for the sake of the stress calculations will be taken as equal to its maximum possible value. The motor current i_d will be also assumed constant ($i_d=I_d$) for the analytical analysis. This assumption is equivalent to considering the branch where the motor is placed to be a current source.

2.1 Analytical Investigation of the SRDCL Converter; Maximum Power Control Scheme

Operation of the SRDCL converter when maximum power transfer methodology is used, can be analyzed in 5 different modes as explained below.

Mode 1: Mode 1 has two different intervals. To begin, it can be assumed that the capacitor has an initial voltage of V_{c0} and converter output voltage is $v_d = V_d$. During this interval the switch is assumed to be off and therefore the resonant current $i_s=0$. The constant current I_d flows through the capacitor to charge it until the voltage across the switch is equal V_{swt} so that it can be turned on. The equation for this interval is:

$$v_c(t)=\frac{-I_d}{C_0}t+V_{c0} \qquad (1)$$

This interval ends when $v_{sw}(t)=V_{swt}$, i.e. $v_c(t)= V_d-V_{swt}$.

Since the voltage spanning the switch has the required positive value the device can be turned on and the second interval begins. The elements L_0 and C_0 establish a resonant circuit and the following equations for the capacitor voltage and inductor current define the resonant operation.

$$v_c(t) = V_d - V_{swt}\cos\omega_0 t - Z_0 I_d \sin\omega_0 t \qquad (2)$$

$$i_s(t)= I_d + \frac{V_{swt}}{Z_0}\sin\omega_0 t - I_d\cos\omega_0 t \qquad (3)$$

where Z_0 and ω_0 are given with following equations.

$$Z_0=\sqrt{L_0/C_0} \qquad (4a)$$

$$\omega_0=\frac{1}{\sqrt{L_0C_0}} \qquad (4b)$$

When the resonant current i_s reaches zero, the switch is turned off. At this moment, the capacitor voltage is given as $v_c(tf)= V_d + V_{swt}$. Due to the assumption of periodicity, this voltage is also equal to the initial value of the capacitor voltage of the first interval. Mode 1 continues as long as there is demand for more motor current. If, at the end of a resonant pulse, the motor current is larger than the reference current, then the switches that will provide the most negative converter output voltage must be turned on so that the motor current then tends to decrease. This requirement is accomplished in Mode 2.

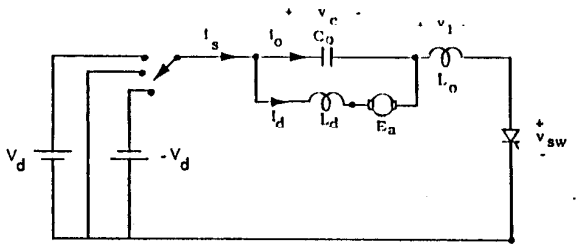

Fig.3 The monophase equivalent circuit of the SRDCL Converter

Mode 2: In this mode the capacitor is charged from (V_d+V_{swt}) to (V_d-V_{swt}). The charging equation is:

$$v_c(t) = \frac{-I_d}{C_0}t + V_d + V_{swt} \qquad (5)$$

Mode 2 ends when $v_{sw}(t)=V_{swt}$.

Mode 3: In this mode the converter output voltage is negative, $v_d(t)= -V_d$. A new resonant pulse is produced. The equations for this mode are given below.

$$v_c(t) = -V_d \quad V_{swt}\cos\omega_0 t \quad Z_0 I_d \sin\omega_0 t \qquad (6)$$

$$i_s(t)= I_d + \frac{V_{swt}}{Z_0}\sin w_0 t - I_d\cos w_0 t \qquad (7)$$

When the resonant current goes to zero a new charging period follows as long as there is demand for negative output voltage, i.e. less motor current.

The equation for the charging of the capacitor from $-V_d+V_{swt}$ to $-V_d-V_{swt}$ is:

$$v_c(t) = \frac{-I_d}{C_0}t - V_d + V_{swt} \qquad (8)$$

If more current is needed, the most positive output voltage available is required. Therefore after the resonant pulse reaches zero Mode 4 starts.

Mode 4: If one considers that the capacitor voltage at the end of the resonant pulse in Mode 2 is equal to $-V_d + V_{swt}$ and v_d will be equal to V_d in the next mode, the switch voltage is found as

$$v_{sw}(t) = 2V_d - V_{swt} \qquad (9)$$

This voltage is always greater than V_{swt}, and, therefore, the next pair of switches can be turned on without any charging period. During this mode, $v_d(t)=V_d$ and $V_{c0}=-V_d+V_{swt}$. The equations for this resonant mode are given below.

$$v_c(t) = V_d-(2V_d-V_{swt})\cos\omega_0 t - Z_0 I_d \sin\omega_0 t \qquad (10)$$

$$i_s(t)= I_d + \frac{2V_d-V_{swt}}{Z_0}\sin\omega_0 t - I_d\cos\omega_0 t \qquad (11)$$

As Eqs. 10 and 11 clearly show, the maximum values of the capacitor voltage and the resonant current are considerably greater than those in the other modes. At the end of the first current pulse, the capacitor voltage is equal to $3V_d - V_{swt}$ and the switch voltage is $-2V_d + V_{swt}$. This requires another charging period which is designated as Mode 5.

Mode 5: During this charging period, the equation is:

$$v_c(t) = \frac{-I_d}{C_0}t + 2V_d - V_{swt} \qquad (12)$$

This mode ends when $v_{sw}(t)= V_{swt}$ and $v_c(t)=V_d-V_{swt}$.

When Mode 5 is over, Mode 1 starts again and continues as long as is necessary as explained before.

Figure 4 shows the operation waveforms of the SRDCL converter based on the single phase equivalent circuit. Waveforms are for the case of $V_d=2V_{swt}$.

2.2 Analytical Investigation of the SRDCL Converter: Adjacent States Vector Control Scheme

Operation of the SRDCL converter for this control method is analyzed below in the same way with the maximum power transfer control method. There are 6 different modes of operation.

Mode 1: Mode 1 for this method is exactly the same as Mode 1 of the previous method. The same equations and explanations are valid. If a negative voltage is required to pull down the motor current, a null state must be utilized first following a charging period.

Mode 2: At this mode the capacitor voltage is charged from V_d+V_{swt} to $-V_{swt}$ so that the switch would have the required positive voltage at the end. The equation is:

$$v_c(t) = \frac{-I_d}{C_0} t + V_d + V_{swt} \qquad (13)$$

Mode 3: This null state mode starts with $V_{c0}=-V_{swt}$ and $v_d(t)=0$. The following equations define the operation.

$$v_c(t) = -V_{swt} \cos \omega_0 t - Z_0 I_d \sin \omega_0 t \qquad (14)$$

$$i_s(t)= I_d + \frac{V_{swt}}{Z_0} \sin \omega_0 t - I_d \cos \omega_0 t \qquad (15)$$

This mode ends when the first current pulse reaches zero. At this moment the capacitor voltage is V_{swt} and a charging period is required before applying the most negative voltage.

Mode 4: The capacitor charges from V_{swt} to $-V_d-V_{swt}$ with the following equation,

$$v_c(t) = \frac{-I_d}{C_0} t + V_{swt} \qquad (16)$$

Mode 5: For this mode $v_d(t)=-V_d$ and $V_{c0}= -V_d-V_{swt}$. The same equations as those of Mode 3 for the previous control scheme are valid. At the end of the current pulse the capacitor voltage is $-V_d+V_{swt}$ and a charging period has to follow. The capacitor is charged from $-V_d + V_{swt}$ to $-V_d - V_{swt}$ with the following equation.

$$v_c(t) = \frac{-I_d}{C_0} t - V_d + V_{swt} \qquad (17)$$

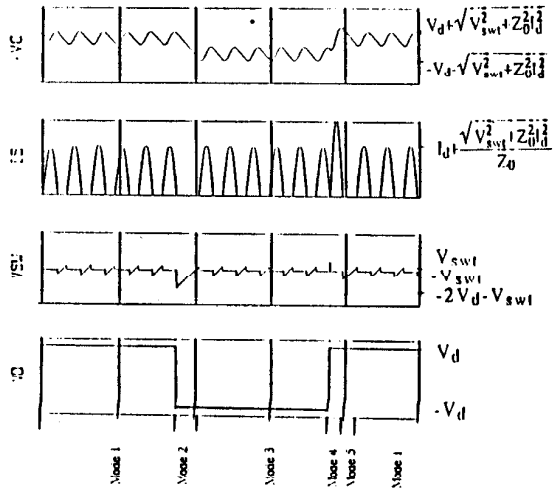

Fig.4 Waveforms of the SRDCL Converter with Maximum Power Transfer Method

Mode 5 continues until a positive voltage demand arises.

Mode 6: If more current is required a positive voltage has to be applied to the resonant link. Therefore the converter has to go through a charging period and a null state before a positive voltage is applied. Mode 6 starts when the resonant current dies out in Mode 5. The initial condition for the capacitor voltage is $-V_d+V_{swt}$, and therefore the switch voltage is V_d-V_{swt} since $v_d(t)$ is zero. This switch voltage is greater than or equal to V_{swt} if V_{swt} is less than or equal to 0.5 V_d and in this case no charging period is required. The equations during this null state period are:

$$v_c(t) = -(V_d-V_{swt}) \cos \omega_0 t - Z_0 I_d \sin \omega_0 t \qquad (18)$$

$$i_s(t)= I_d + \frac{V_{swt}}{Z_0} \sin \omega_0 t - I_d \cos \omega_0 t \qquad (19)$$

At the end of the resonant pulse, $v_c(t)=V_d-V_{swt}$ and, $v_{sw}(t)=V_{swt}$ since the converter output voltage is going to be V_d for the next period. This means that the switch can be turned on again without any need for a charging period. When the switch is turned on, a resonant pulse is produced and Mode 1 again begins.

If V_{swt} is greater than 0.5 V_d, then a charging period is required before the null state resonant pulse. In that case the capacitor has to be charged from $-V_d+V_{swt}$ to $-V_{swt}$. Then a null state would follow with the following equations.

$$v_c(t) = -V_{swt} \cos \omega_0 t - Z_0 I_d \sin \omega_0 t \qquad (20)$$

$$i_s(t)= I_d + \frac{V_{swt}}{Z_0} \sin \omega_0 t - I_d \cos \omega_0 t \qquad (21)$$

At the end of the resonant pulse the capacitor voltage and the switch voltage would be V_{swt} and V_d-V_{swt} respectively. Since this switch voltage is smaller than V_{swt} for this case, another charging period would be required. In such a charging period the capacitor would be charged from V_{swt} to V_d-V_{swt} after which point Mode 1 operation would again start. Figure 5 shows the waveforms of the SRDCL converter for the adjacent state vector control method.

2.3 Voltage and Current Stresses of the SRDCL Converter

The voltage and current stresses for the converter in each mode are summarized below for the both control methods. In order to realize equations in terms of V_d, V_{swt} = K V_d was used. Following table shows the stresses when the maximum power transfer method is used.

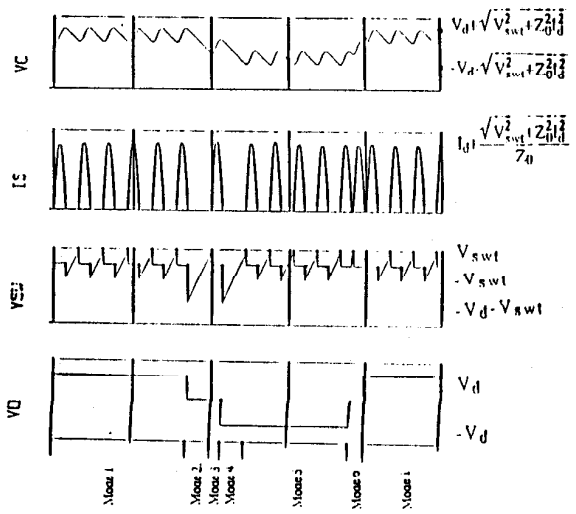

Fig.5 Waveforms of the SRDCL Converter with Adjacent States Vector Control Method

Mode	$\lvert v_{cmax}\rvert$	$\lvert i_{smax}\rvert$
1	$V_d+\sqrt{K^2V_d^2+Z_0^2I_d^2}$	$I_d+\sqrt{K^2V_d^2+Z_0^2I_d^2}/Z_0$
2	$(1+K)\,V_d$	0
3	Same as Mode 1	Same as Mode 1
4	$V_d+\sqrt{(2-K)^2V_d^2+Z_0^2I_d^2}$	$I_d+\sqrt{(2-K)^2V_d^2+Z_0^2I_d^2}/Z_0$
5	$(2-K)\,V_d$	0

Table 1
Voltage and current stresses in each mode for the maximum power control scheme

As can be seen from this table the largest voltage and current stresses occur during Mode 4. In order to have normalized stress equations the voltage equation can be divided by V_d, and the current equation can be divided by I_d. This gives the normalized maximum voltage (V_{sn}) and current (I_{sn}) values for this method.

$$V_{sn} = 1 + \sqrt{(2-K)^2 + Z_0^2 I_d^2 / V_d^2} \qquad (22)$$

$$I_{sn} = 1 + \sqrt{1+(2-K)^2 V_d^2 / (Z_0^2 I_d^2)} \qquad (23)$$

Figure 6 shows the variation of these stresses for K values between 0.1 and 0.9 when $Z_0 = 10\ \Omega$, $I_d = 10$ A and $V_d = 200$ V. The figure shows that the peak capacitor voltage may reach 2.9 p.u. for K=0.1 while the peak current value is about 4.9 p.u for the same K value. For a reasonable K value of 0.5, these values are 2.58 p.u. and 4.16 p.u. respectively.

Table 2 shows the voltage and current stresses when the adjacent states method is used.

Mode	$\lvert v_{cmax}\rvert$	$\lvert i_{smax}\rvert$
1	$V_d+\sqrt{K^2V_d^2+Z_0^2I_d^2}$	$I_d+\sqrt{K^2V_d^2+Z_0^2I_d^2}/Z_0$
2	$(1+K)\,V_d$	0
3	$\sqrt{K^2V_d^2+Z_0^2I_d^2}$	Same as Mode 1
4	Same as Mode 2	0
5	Same as Mode 1	Same as Mode 1
6a	$V_d+\sqrt{(1-K)^2V_d^2+Z_0^2I_d^2}$	$I_d+\sqrt{(1-K)^2V_d^2+Z_0^2I_d^2}/Z_0$
6b	Same as Mode 3	Same as Mode 1

Table 2
Voltage and current stresses in each mode for the adjacent state vector control scheme

Mode 6a and 6b correspond to the cases where $K \le 0.5$ and $K > 0.5$ respectively. The maximum stresses in this method occur in Mode 2 and their normalized values are given below.

$$V_{sn} = 1 + \sqrt{K^2 + Z_0^2 I_d^2 / V_d^2} \qquad (24)$$

$$I_{sn} = 1 + \sqrt{1+K^2 V_d^2 / (Z_0^2 I_d^2)} \qquad (25)$$

Figure 7 shows the variation of these values for K is between 0.1 and 0.9 with the same Z_0, I_d and V_d values as before. For this method, the peak capacitor voltage and the peak resonant current are about 2 p.u. and 3.05 p.u. respectively. Current stress grows as K increases and the biggest current stress occurs for K=0.9. The voltage stress is almost constant for all values of K. It should be noted that for K=0.1 the increase in the peak current value in Mode 6 is not significantly greater than that in Mode 2 (2.06 p.u. and 2.02 p.u. respectively).

When Fig. 6 and Fig. 7 are compared it is easily seen that by using the adjacent state method instead of maximum power transfer method the voltage and current stresses can be decreased drastically. The worst maximum capacitor voltage for the adjacent states method is still lower than the lowest value of the maximum voltage stress of the previous

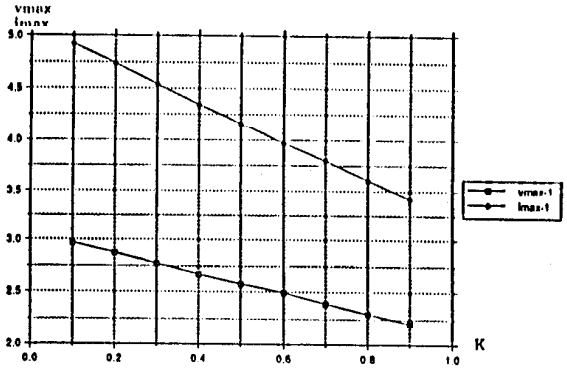

Fig.6 Voltage and current stresses for Maximum Power Transfer Method

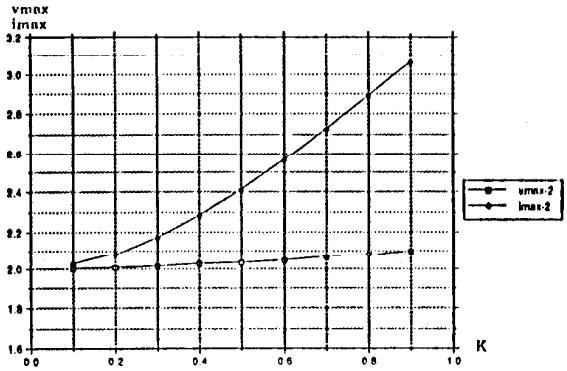

Fig.7 Voltage and current stresses for Adjacent States Vector Control Method

method. The same is true for the current values. The biggest peak current value for the adjacent states method is 3.06 p.u. which is lower than the smallest peak current value of the other method, namely 3.4 p.u.

3 Simulation Results

The SRDCL Converter driving a dc motor was simulated by using ACSL simulation software. The parameters of the dc motor are given as:

Rated power: 10 HP
L_a=4.3 mH
R_a=0.57 Ω
J=0.0881 kgm/rad/sec^2
B=0.02 kgm/rad/sec
ω_{rated}=1750 rpm = 183.26 rad/sec
K_v =1.17 V/sec
K_t =1.17 Nm/A
Smoothing inductor=40 mH

Figures 8 and 9 show the converter output voltage and the resonant current waveforms for the maximum power control method and adjacent states method respectively. This waveforms verify that lower peak current stresses occur when the latter explained method is utilized. Figure 10 is given to show the unity input power factor feature of the converter. In the figure, the input voltage and unfiltered input current for one phase are shown.

4 Experimental Results

The system that is shown in Fig. 1 has been constructed in our laboratory with the following component values and ratings:

C_0 =0.9 μF

L_0 =60 μH

L_d =30 mH

Input voltage: 115 V, 60 Hz

C_f =50 μF

The DC motor utilized is a 3/4 HP permanent magnet machine. IR85RDT GTO's were used as the switching devices. The value of V_{swt} was set to 75 V which is about $0.46V_d$. The system was controlled by a Motorola 56000 DSP microprocessor.

Control of the motor was achieved by two different methods. The first method corresponds to the maximum power transfer control method which was described earlier. In this method, depending on the current error, the switches that would yield the most positive or most negative converter output voltage are turned on. Figure 11 shows the capacitor voltage and resonant current waveforms for this control method. As explained before, when there is a transition from negative capacitor voltage to positive capacitor voltage, there is an irregular current pulse. In this case these pulses have peak values 2.5 times the regular peak values.

The second method implemented was a modified version of the first approach. In this method, when there is a demand for less current, null state is chosen and kept until

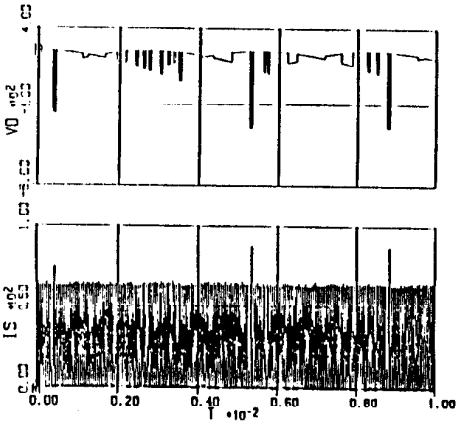

Fig.8 Simulation result: Converter output voltage and resonant current with Maximum Power Transfer Method

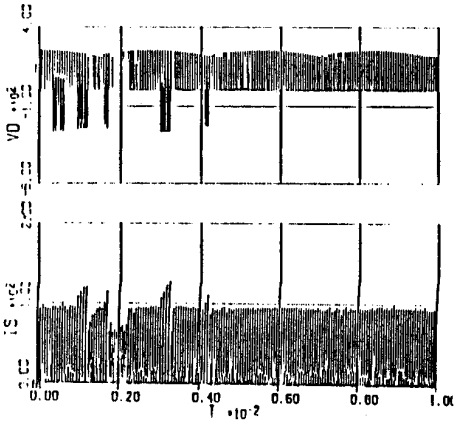

Fig.9 Simulation result: Converter output voltage and resonant current with Adjacent States Vector Control Method

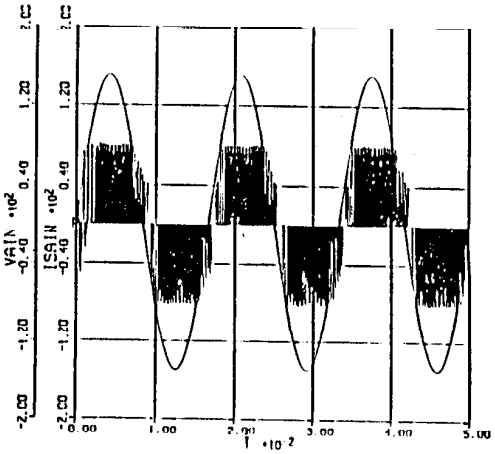

Fig.10 Simulation result: Input voltage and current of the SRDCL Converter

there is demand for more current. There is no negative voltage mode. Figure 12 shows the capacitor voltage and resonant current for this method. As expected, the capacitor voltage stress is less and the resonant current pulses are almost uniform due to the fact that the capacitor does not need to charge from a negative value to a positive value.

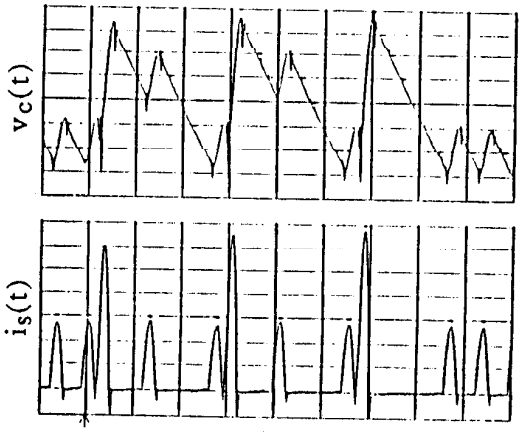

Fig.11 Experimental waveforms with Maximum Power Transfer Method (100 V/div, 5 A/div)

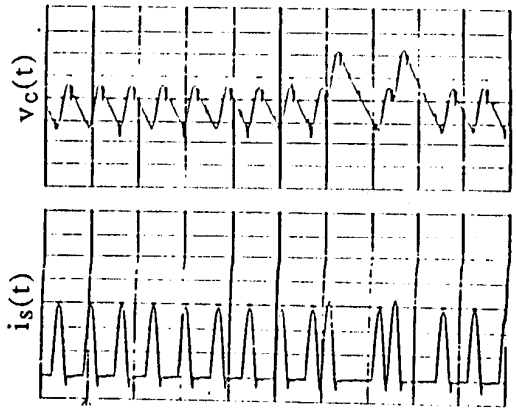

Fig.12 Experimental waveforms with modified strategy (100 V/div, 5 A/div)

Figure 13 shows the capacitor voltage, motor speed, motor current and resonant current during the start-up process when the second method is used. It is clear that a substantial improvement in performance has been obtained

5 Conclusion

In this paper, an improved control algorithm for a series resonant dc current link converter was described and applied as a dc motor drive. The special attention was paid to component stresses. Stress calculations for two different methods, maximum power transfer and adjacent state vector control methods were presented and compared. It was shown that, by using the latter explained method, component stresses can be reduced significantly by only modifying the control scheme, at no extra cost. This makes the SRDCL converter, which already has low switching losses even at high frequencies due to the zero current switching, more attractive and makes it possible to have lower rated devices and resonant components.

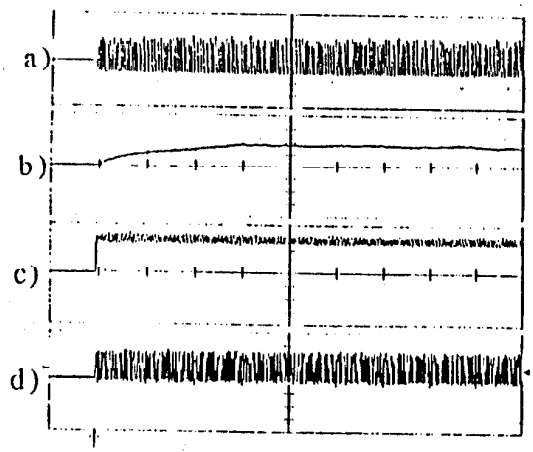

Fig.13 Starting waveforms of DC motor with modified strategy. a) v_d,100 V/div,b) ω_r,250 rpm/div, c) i_d, 5A/div, d) is, 5A/div

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